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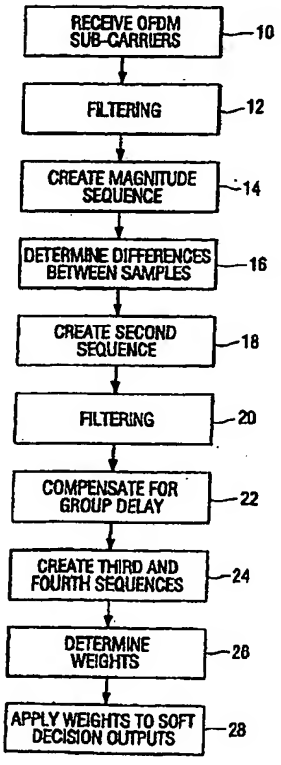
(54) Title: ADAPTIVE WEIGHTING METHOD FOR ORTHOGONAL FREQUENCY DIVISION MULTIPLEXED SOFT SYMBOLS USING CHANNEL STATE INFORMATION ESTIMATES

(57) Abstract

A method for weighting orthogonal frequency division multiplexed soft symbols is provided, including the steps of receiving a plurality of sub–carriers modulated by digital information and filtering the sub–carriers to produce complex soft decision outputs. The magnitudes of the soft decision outputs are used to creating a first sequence of data. The differences between successive samples in the first sequence are used to create a second sequence of data. The first and second sequences are filtered and compensated for differences in effective group delay of the first and second sequences to produce third and fourth sequences. Then these third and fourth sequences are used to determine a plurality of weights and the weights are applied to the complex soft decision outputs. A receiver which incorporates the method is also disclosed.



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ADAPTIVE WEIGHTING METHOD FOR ORTHOGONAL FREQUENCY DIVISION MULTIPLEXED SOFT SYMBOLS USING CHANNEL STATE INFORMATION ESTIMATES

BACKGROUND OF THE INVENTION

This invention relates to signal processing, and more particularly, to signal processing techniques for use in Digital Audio Broadcasting (DAB) systems.

Digital Audio Broadcasting is a medium for providing digital-quality audio, superior to existing analog broadcasting formats. Both AM and FM DAB signals can be transmitted in a hybrid format where the digitally modulated signal coexists with the currently broadcast analog AM or FM signal, or in an all-digital format without an analog signal. In-band-on-channel (IBOC) DAB systems require no new spectral allocations because each DAB signal is simultaneously transmitted within the same spectral mask of an existing AM or FM channel allocation. IBOC promotes economy of spectrum while enabling broadcasters to supply digital quality audio to their present base of listeners. Several IBOC DAB approaches have been suggested. One such approach, set forth in U. S. Patent No. 5,588,022, presents a method for simultaneously broadcasting analog and digital signals in a standard AM broadcasting channel. Using this approach, an amplitude-modulated radio frequency signal having a first frequency spectrum is broadcast. The amplitude-modulated radio frequency signal includes a first carrier modulated by an analog program signal. Simultaneously, a plurality of digitally-modulated carrier signals are broadcast within a bandwidth which encompasses the first frequency spectrum. Each digitally-modulated carrier signal is modulated by a portion of a digital program signal. A first group of the digitally-modulated carrier signals lies within the first frequency spectrum and is modulated in quadrature with the first carrier signal. Second and third groups of the digitally-modulated carrier signals lie outside of the first frequency spectrum and are modulated both in-phase and

in-quadrature with the first carrier signal. Multiple carriers are employed by means of orthogonal frequency division multiplexing (OFDM) to bear the communicated information.

FM IBOC broadcasting systems using have been the subject of several United States patents including patents no. 5,465,396; 5,315,583; 5,278,844 and 5,278,826. In addition, a commonly assigned pending patent application for a "Method and System for Simultaneously Broadcasting and Receiving Digital and Analog Signals, by D. Kumar and B. Hunsinger, Serial No. 08/294,140, filed July 1994 discloses an FM IBOC DAB system.

The signals used in Digital Audio Broadcasting are subject to fading and noise (interference). Digital Audio Broadcasting receivers may include Viterbi decoders. Conventional implementations of soft-decision Viterbi decoders rely on constant signal and gaussian noise statistics for (near) optimum decoding. Practically these statistics should be nearly constant over the path memory of the Viterbi decoder, or the span of the interleaver, whichever is greater. An interleaver may be used to yield statistical independence of the fading statistics of the soft symbols over the path memory of the Viterbi decoder after deinterleaving. The remedy for a flat fading channel causing fluctuating signal levels with constant noise is well known. However, there exists a need for a signal processing technique that can address independently varying signal and noise levels.

SUMMARY OF THE INVENTION

A method for weighting orthogonal frequency division multiplexed soft symbols is provided, including the steps of receiving a plurality of sub-carriers modulated by digital information, filtering the sub-carriers to produce complex soft decision outputs, creating a first sequence of the magnitudes of the complex soft decision outputs, determining the differences between successive samples in the first sequence, creating a second sequence of the differences between successive samples in the first sequence, filtering the first and second sequences and compensating for differences in effective group delay of the first and second sequences to produce third and fourth sequences, using the third and fourth sequences to determine a plurality of weights, and applying the plurality of weights to the complex soft decision outputs. The invention also encompasses receivers that incorporate the above method.

BRIEF DESCRIPTION OF THE DRAWINGS

Figure 1 is a diagram showing the steps in the method of this invention; and

Figure 2 is a functional block diagram showing the mapping and processing of bits through a digital audio broadcasting receiver which employs the method of this invention.

DESCRIPTION OF THE PREFERRED EMBODIMENTS

The present invention will be described in terms of the adaptive weighting procedure for coherently detected quadrature phase shift keyed (QPSK) or binary phase shift keyed (BPSK) signals under various fading conditions assuming gaussian noise or interference statistics.

For single carrier, flat fading signal, constant noise or interference, the flat fading channel multiplies the k^{th} interleaved symbol by a fading factor a_k, which is assumed to be constant over the symbol duration. The signal sample output of a matched filter in the receiver can be expressed as:

$$y_k = a_k \cdot x_k + n_k \tag{1}$$

where x_k is the transmitted signal component and n_k is the received noise. The appropriate weighting for that soft decision symbol at the receiver is itself a_k, or more practically the receiver's estimate $\hat{a}_k$. Therefore the soft symbol s_k entering a deinterleaver prior to the Viterbi decoder is presented as:

$$s_k = \hat{a}_k \cdot y_k \tag{2}$$

The symbols immediately surrounding the k^{th} symbol are similarly affected by the relatively slowly varying fading. It is impossible to adequately estimate this fading factor over a single symbol because it is impossible to separate its statistics from the noise. However the fading factor can be estimated statistically using a weighted average over a set of samples with correlated fading constants, specifically the symbols that are temporally close to the k^{th} symbol.

Let the signal term x_k be complex whose real and imaginary components are binary valued at ± 1. For reasonably large SNR, where the noise (or interference) is approximated to be zero mean average white Gaussian noise (AWGN), the fading factor might be estimated as:

$$\begin{aligned} \hat{a}_k &= E\{|y_k|\} \\ &= E\{|a_k \cdot x_k + n_k|\} \cong E\{a_k\} \end{aligned} \tag{3}$$

However, this estimate is not accurate for small SNR since the rectified noise will place a lower limit on $\hat{a}_k$. This limitation would reduce erasure effectiveness when the signal level fades to zero. The signal level for coherent detection and Gaussian noise can be estimated using an exact expression as a function of the second and fourth moments of the magnitude of y_k (i.e. $E\{|y_k|^2\}$ and $E\{|y_k|^4\}$). These moments are obtained through averaging or lowpass filtering of the squared or fourth power of the magnitude of y_k. Assuming a constant AWG noise floor, the weighting factor for the soft decision is equal to the fading factor:

$$\hat{a}_k = \sqrt{2 \cdot (E\{|y|^2\})^2 - E\{|y|^4\}} \tag{4}$$

This equation has been used to successfully estimate the signal level in a mobile satellite receiver.

For single carrier, flat fading signal, and flat fading interference, the difference between the constant signal and the flat fading signal appears as an interference fading factor σ_k. It is assumed that the fading of the signal and interference are approximately independent, and the interference is approximated as AWGN. The Gaussian interference assumption is made in the absence of known precise statistics of the interferer. Furthermore, this assumption becomes less objectionable when we recognize that the Viterbi decoder tends to "Gaussianize" the noise or interference summed over the weighted symbols in its path memories. The soft-decision output of the matched filter can be expressed as:

$$y_k = a_k \cdot x_k + \sigma_k \cdot n_k \tag{5}$$

Dividing both sides by σ_k puts the expression in form familiar to the nonfaded interference case:

$$\frac{y_k}{\sigma_k} = \frac{a_k}{\sigma_k} \cdot x_k + n_k \tag{6}$$

Then the appropriate weight for the soft symbol is identified. The soft symbol entering the deinterleaver prior to the Viterbi decoder is presented as:

$$s_k = \frac{\hat{a}_k}{\hat{\sigma}_k^2} \cdot y_k \tag{7}$$

The fading factor can be estimated as in equation (4). An additional estimate of the noise variance $\hat{\sigma}_k^2$ is required for this weighting. This estimate can be computed as:

$$\hat{\sigma}_k^2 = E\{|y_k|^2\} - \hat{a}_k^2 \tag{8}$$

For multiple sub-carriers (OFDM), selective fading signal, and selective fading interference, the method of the present invention can be described as follows. OFDM sub-carriers comprise a set of orthogonally spaced (in frequency) sub-carriers, each modulated synchronously at the symbol rate. QPSK sub-carrier modulation is assumed here, although other linear modulation techniques such as MPSK or QAM can be accommodated with modification. OFDM has been shown to be tolerant of multipath fading since the fading over any individual sub-carrier bandwidth is considered to be flat (flat fading requires no sub-carrier equalization). In a severe selective fading instance, a portion of the OFDM sub-carriers may be lost in a null while the remaining sub-carriers are recoverable. In contrast, wideband single carrier modulation can experience the same spectral null where significant degradation can occur over all the bits if equalization is not employed.

Each OFDM sub-carrier can be treated as an independent channel with flat fading signal and interference; however, this is not optimum. The correlation of the signal and noise fading across sub-carriers nearby in frequency can be exploited to improve performance. This can be accomplished through filtering the statistics across both time and frequency surrounding any particular sub-carrier. The k^{th} symbol of the n^{th} sub-carrier is weighted as:

$$s_{k,n} = \frac{\hat{a}_{k,n}}{\hat{\sigma}^2_{k,n}} \cdot y_{k,n} \tag{9}$$

where $y_{k,n}$ is a complex soft symbol resulting from QPSK demodulation of the k^{th} symbol of the n^{th} sub-carrier. The only difference between equations (9) and (7) is the exploitation of the correlation of adjacent sub-carrier statistics in equation (9). The fading factors can be estimated after 2-dimensional smoothing (filtering across time and sub-carriers) of the following values:

$$\begin{aligned} u_{k,n} &= E\{|y_{k,n}|^2\} \\ v_{k,n} &= E\{|y_{k,n}|^4\} \end{aligned} \tag{10}$$

Compute the symbol weighting factor for coherently detected QPSK as:

$$w_{k,n} = \frac{\sqrt[4]{2 \cdot u^2_{k,n} - v_{k,n}}}{u_{k,n} - \sqrt{2 \cdot u^2_{k,n} - v_{k,n}}} \tag{11}$$

where $w_{k,n}$ is the weight to be used for the soft symbol prior to deinterleaving.

The adaptive weighting procedure for differentially detected QPSK signals under various fading conditions will now be described. These results can be extended to BPSK with minor modification.

The goal of the present invention is to maximize the resulting SNR of the weighted and combined symbols after differential detection into the Viterbi decoder. This minimizes the probability of bit error at the output of the Viterbi decoder. The absolute optimization of the weighting of the soft decision statistic is a function of the particular Viterbi decoder. Therefore the answer is not necessarily unique. Furthermore, since the Viterbi decoder tends to "Gaussianize" the path metrics over the path memories, then this technique tends to maximize the SNR over the path memory.

A similar expression for the weighting factor can be derived assuming Gaussian noise into a differential QPSK detector resulting in non-Gaussian statistics at the output. The fading factor can be computed as a function of the statistics of the output of the differential detector where we define the soft decision of the form:

$$S = (a + n_1) \cdot (a \cdot e^{j\phi} + n_2) \tag{12}$$

where ϕ denotes the phase information imposed between a pair of adjacent symbols in the differential encoding, and n are the independent noise samples. The fading factor a of the adjacent symbols is assumed to be approximately equal. The signal to noise ratio after differential detection is easily computed to be:

$$SNR = \frac{a^4}{2 \cdot a^2 \cdot \sigma^2 + \sigma^4} \tag{13}$$

The ideal weighting factor for the post-differentially detected symbols is therefore:

$$w = \frac{a^2}{2 \cdot a^2 \cdot \sigma^2 + \sigma^4} \tag{14}$$

The first differential approach described here uses statistical estimates of the second and fourth moments of the differentially detected symbols to form the weighting factor. The second and fourth moments are described by the following previously known relationships.

$$\begin{aligned} E\{|S|^2\} &= (a^2 + \sigma^2)^2 \\ E\{|S|^4\} &= (a^4 + 4 \cdot a^2 \cdot \sigma^2 + 2 \cdot \sigma^4)^2 \end{aligned} \tag{15}$$

Then the fading factor can be estimated as :

$$\hat{a}_k = \sqrt[4]{2 \cdot E\{|S_k|^2\} - \sqrt{E\{|S_k|^4\}}} \tag{16}$$

and the noise can be estimated as:

$$\hat{\sigma}^2{}_k = \sqrt{E\{|S_k|^2\}} - \hat{a}_k^2 \tag{17}$$

The estimates of equations (16) and (17) are inserted into equation (14) to obtain the weight.

Simulations were performed using adaptive weighting as described in equations (14), (16), and (17). Although long-term estimates without fading yielded good results, a compromise must be reached between long filter time constants for accurate estimation versus short filter time constants needed to track varying statistics due to fading.

In the Digital Audio Broadcasting (DAB) simulation, the OFDM symbol rate of 689.0625 Hz was chosen with a fading bandwidth of 13 Hz. Then the reciprocal of the fading bandwidth is about 53 symbols in this case. A filter time constant of 16 symbols was chosen since this time constant must be small compared to the fading time. Unfortunately, the statistical estimation errors over this short filter time yielded poor performance results for the adaptive weighting compared to what would be possible with perfect statistical estimation. Even reducing the fading bandwidth down to 3 Hz and increasing the filter time constant to 64 samples left a significant loss.

Equations (16) and (17) reveal that, in effect, quantities raised to the fourth power are subtracted to yield smaller numbers. This situation is most pronounced when the signal and noise powers are approximately equal, resulting in large estimation errors. Simulation results support this observation. Therefore another estimator is sought that does not rely upon subtraction fourth order statistics. The desired estimation technique should be designed to accommodate a fading bandwidth of up to 13 Hz for maximum vehicle speeds in the FM band around 100 MHz.

The optimum soft-symbol weight to be applied before differential detection of QPSK can be described as a function of time (k index) and OFDM sub-carrier (n index). Similar to equation (14), this weight is:

$$w_{k,n} = \frac{a_{k,n}}{\sqrt{2a_{k,n}^2\sigma_{k,n}^2 + \sigma_{k,n}^4}} \tag{18}$$

where $a_{k,n}$ is the fading coefficient of the k^{th} symbol for the n^{th} sub-carrier. and $\sigma_{k,n}$ is the corresponding standard deviation of the noise or interference, both prior to differential detection. Notice that the weight of equation (18) is the square root of equation (14). This is a result of the reasonable assumption that the weight changes slowly over the symbol-pair

time used in the differential detection. In effect, the differential detection squares the predetection weight of equation (18), which would result in equation (14). A method for improving the statistical estimates of equation (18) is sought.

Practical methods for estimating CSI and weights using pre-differentially detected soft-symbols and weight also applied to the soft decision symbol prior to differential detection will now be discussed.

For moderate to high SNR, the weight of equation (18) can be conveniently approximated by:

$$\lim_{SNR \to \infty} w_k = \frac{1}{\sqrt{2\sigma_k^2}}, \tag{19}$$

where simple statistical measurements were used to estimate σ^2. However, simulation confirmed that this weight estimate performed poorly during times when the SNR was very low due to fading interference. For example, the optimum weight would have suppressed the noisy samples more than the high SNR approximation to the weight. Therefore, another approximation was sought which would estimate CSI statistics over a large SNR range. Furthermore the estimate should not be sensitive to a gaussian noise or interference assumption, and should be estimated with sufficient accuracy in a time (filter time constant) significantly less than the reciprocal of the fading bandwidth.

A simple and robust estimation technique evolved after simulation and some experimentation. This estimation technique approximates the previously-defined weight expressions, but uses lower-order statistical approximations. This technique is described in the following 4 steps.

1. Create a sequence $v_{k,n}$ for each QPSK sub-carrier consisting of the magnitudes of the complex soft decision outputs $s_{k,n}$ from the matched filter for the nth sub-carrier.

$$v_{k,n} = |s_{k,n}| \tag{20}$$

2. Create a sequence $d_{k,n}$ consisting of the differences of successive time samples of $v_{k,n}$.

$$d_{k,n} = v_{k,n} - v_{k-1,n} \tag{21}$$

3. Filter the sequences $v_{k,n}$ and $d_{k,n}$ using second-order digital IIR filters, then compensate for any differences in effective group delay to yield sequences $filtv_{k,n}$ and

$filtd_{k,n}$. The time constant for the $filtv_{k,n}$ filter should be somewhat smaller than the reciprocal of the fading bandwidth, while the time constant for the $filtd_{k,n}$ filter can be somewhat larger. These sequences are representative (approximately proportional) of the local mean and standard deviation of the sequence $v_{k,n}$.

4. The sequence of weights for the soft decisions for each sub-carrier to be applied prior to differential detection is defined as

$$w_{k,n} = \frac{1}{filtd_{k,n} \cdot \left(1 + \left(\frac{filtd_{k,n}}{filtv_{k,n} - filtd_{k,n}}\right)^4\right)} \tag{22}$$

To prevent numerical overflow, check to ensure that $filtv_{k,n} > 1.5 \cdot filtd_{k,n}$ in equation (22); otherwise, set the weight to zero. Simulation results verified that this weight yields good performance under a variety of channel impairments with fading and interference.

The values of $filtd_{k,n}$ and $filtv_{k,n}$ are estimated using filtering techniques described next. Filtering is performed first for each sub-carrier at the k^{th} symbol instant in time. Then the rows of $filtd_{k,n}$ and $filtv_{k,n}$ are simply updated across the N sub-carriers. Equation (23) filters the sequences $v_{k,n}$ with a time delay of approximately 16 symbols, and equation (24) filters the sequences $d_{k,n}$ with a time delay of approximately 64 symbols. Both filters have a zero frequency gain of nearly unity.

$$subv_{k,n} = \frac{960 \cdot subv_{k-1,n} - 451 \cdot subv_{k-2,n} + 3 \cdot v_{k,n}}{512} \tag{23}$$

$$subd_{k,n} = \frac{16128 \cdot subd_{k-1,n} - 7939 \cdot subd_{k-2,n} + 3 \cdot d_{k,n}}{8192} \tag{24}$$

Additional filtering is performed across the N sub-carriers. Smoothing the estimates across the N sub-carriers requires 3 passes of a simple IIR filter. The first pass sets the appropriate initial condition of the filter, but does not update the estimates. The direction of the second pass is reversed from the first, while the third pass is reversed again. This results in an approximately symmetric (linear phase) filter characteristic which is desirable for providing the estimates on the center carrier. Although it is impossible to provide this symmetric filtering for the sub-carriers at each end of the band, the impulse response "tails" are folded back into the active sub-carriers.

The first pass across the sub-carriers sets the initial values of $filtv_{N-1}$ and $filtd_{N-1}$ without replacing the time-filtered values for each sub-carrier. The time index k is ignored here since it is understood that the filtering over the sub-carriers is performed over each k^{th} OFDM symbol.

$$\begin{aligned} filtv_{N-1} &\Leftarrow (1-\beta)\cdot filtv_{N-1} + \beta \cdot subv_n \ ; \\ filtd_{N-1} &\Leftarrow (1-\beta)\cdot filtd_{N-1} + \beta \cdot subd_n \\ n &= 0,1,\ldots N-1 \end{aligned} \tag{25}$$

The second pass smoothes the values across the filtered estimates for each sub-carrier, *subv* and *subd*.

$$\begin{aligned} filtv_n &\Leftarrow (1-\beta)\cdot filtv_{n+1} + \beta \cdot subv_n ; \\ filtd_n &\Leftarrow (1-\beta)\cdot filtd_{n+1} + \beta \cdot subd_n ; \\ n &= N-2, N-3, \ldots 0 \end{aligned} \tag{26}$$

The third pass smoothes the frequency values again to achieve a nearly symmetrical impulse response (except for the sub-carriers near the endpoints).

$$\begin{aligned} filtv_n &\Leftarrow (1-\beta)\cdot filtv_{n-1} + \beta \cdot filtv_n ; \\ filtd_n &\Leftarrow (1-\beta)\cdot filtd_{n-1} + \beta \cdot filtd_n ; \\ n &= 1,2,\ldots N-1. \end{aligned} \tag{27}$$

The resulting filtered values for *filtv* and *filtd* computed in equations (26) and (27) are used in equation (22) at each OFDM symbol time to yield the appropriate weight for each soft symbol prior to differential detection, but after matched filtering, in the receiver.

Figure 1 is a diagram showing the steps of the method of this invention. Block 10 shows the step of receiving a plurality of sub-carriers modulated by digital information. The sub-carriers are filtered as shown in block 12 to produce complex soft decision outputs. These outputs are used in block 14 to create a first sequence of the magnitudes of the complex soft decision outputs. Block 16 shows that the differences between successive samples in the first sequence are determined. A second sequence of the differences between successive samples in the first sequence is created as shown in block 18. The first and second sequences are filtered as in block 20 and compensated for differences in effective group delay of the first and second sequences as in block 22 to produce third and fourth sequences in block 24. The third and fourth sequences are used to determine a plurality of weights as shown in block 26, and the plurality of weights are applied to the complex soft decision outputs as shown in block 28.

The above discussion relates to differentially detected QPSK. The derivation for the weight using coherent detection of QPSK or (BPSK) would be similar to the technique discussed above for differentially detected QPSK. The only modification should be in the expression presented in equation (22). It can be predicted that the weight should be of the form :

$$w_{k,n} = \frac{filtv_{k,n} - c \cdot filtd_{k,n}}{filtd_{k,n}^{\ 2}} \tag{27}$$

where c is a constant to be defined by empirical methods.

Techniques for estimating the optimum soft-decision weight for QPSK symbols prior to Viterbi decoding were described. These techniques apply to coherent and differential detection of single or multi-carrier (OFDM) QPSK, with and without multipath fading of the signal of interest or the interferer. The fading cases can necessitate a compromise between accurate CSI estimation and agility of the CSI to track the fading signal or noise components.

The application of soft-decision weighting for an OFDM system with independently faded signal and noise (interference) can improve BER performance. This weight is applied to the soft symbols prior to deinterleaving and Viterbi decoding. The optimal weight for each soft symbol over time and sub-carriers is estimated through an expression using filtered statistical estimates of channel state information (CSI) about the signal and noise components of the received symbols. The above CSI estimation and weighting techniques can apply to both coherent and differentially detected symbols, and to single and multi-carrier (OFDM) modulation in the presence of multipath fading and colored noise statistics.

Figure 2 is a functional block diagram showing the mapping and processing of bits through a of a portion of an FM receiver that operates in accordance with the method of the invention. A plurality of OFDM carriers 30 are received and converted to bit streams on lines 32 by receiver circuit 34. Circuit 34 includes a digitizer, carrier synchronization, symbol synchronization, and matched filters all operating in accordance with well known techniques to produce the bit streams on lines 36. Block 38 represents the channel state estimates and weighting processes performed according to Figure 1 to produce weighted bit streams on lines 40. Block 42 shows that the bit streams are deallocated from the carriers and delivered to a deinterleaver 44. The output of the deinterleaver is multiplexed to a single bit

stream and passed to a Viterbi decoder 46 decodes the single bit stream. Soft-decision Viterbi decoding with (near) optimum soft-decision weighting for maximum ratio combining (MRC) for differentially detected QPSK sub-carrier symbols is employed to minimize losses over the channel. The output of the Viterbi decoder is subject to additional signal processing, which is not part of the present invention, and passed to output stages, as illustrated by block 48, to produce the desired output from the receiver.

While the present invention has been described in terms of what are at present believed to be its preferred embodiments, it will be apparent to those skilled in the art that various changes may be made to the embodiments described above without departing from the scope of the invention as set forth in the following claims.

What is claimed is:

1. A method for weighting orthogonal frequency division multiplexed soft symbols, said method comprising the steps of:

receiving a plurality of sub-carriers modulated by digital information;

filtering the sub-carriers to produce complex soft decision outputs;

creating a first sequence of the magnitudes of said complex soft decision outputs;

determining the differences between successive samples in said first sequence;

creating a second sequence of the differences between successive samples in said first sequence;

filtering said first and second sequences to produce third and fourth sequences;

using said third and fourth sequences to determine a plurality of weights; and

applying said plurality of weights to said complex soft decision outputs.

2. The method of claim 1, wherein said sub-carriers are modulated using quadrature phase shift keying.

3. The method of claim 1, wherein said second and third sequences are filtered using a second order digital infinite impulse response filter.

4. The method of claim 1, wherein:

said third sequence has a time constant smaller than the reciprocal of the fading bandwidth of plurality of sub-carriers; and

said fourth sequence has a time constant larger than the reciprocal of the fading bandwidth of plurality of sub-carriers.

5. The method of claim 1, wherein the step of filtering said first and second sequences further comprises the step of compensating for differences in effective group delay of said first and second sequences to produce third and fourth sequences.

6. The method of claim 1, wherein said plurality of weights are determined using the formula:

$$w_{k,n} = \frac{1}{filtd_{k,n} \cdot \left(1 + \left(\frac{filtd_{k,n}}{filtv_{k,n} - filtd_{k,n}}\right)^4\right)}$$

wherein $w_{k,n}$ represents said plurality of weights, $filtv_{k,n}$ represents said third sequence, and $filtd_{k,n}$ represents said fourth sequence.

7. The method of claim 6, wherein if $filtv_{k,n} > 1.5 \bullet filtd_{k,n}$, then $w_{k,n}$ is set to zero.

8. A radio receiver comprising:

means for receiving a plurality of sub-carriers modulated by digital information;

means for filtering the sub-carriers to produce complex soft decision outputs;

means for creating a first sequence of the magnitudes of said complex soft decision outputs;

means for determining the differences between successive samples in said first sequence;

means for creating a second sequence of the differences between successive samples in said first sequence;

means for filtering said first and second sequences to produce third and fourth sequences;

means for using said third and fourth sequences to determine a plurality of weights; and

means for applying said plurality of weights to said complex soft decision outputs.

9. The receiver of claim 8, wherein said sub-carriers are modulated using quadrature phase shift keying.

10. The receiver of claim 8, wherein said second and third sequences are filtered using a second order digital infinite impulse response filter.

11. The receiver of claim 8, wherein:

said third sequence has a time constant smaller than the reciprocal of the fading bandwidth of plurality of sub-carriers; and

said fourth sequence has a time constant larger than the reciprocal of the fading bandwidth of plurality of sub-carriers.

12. The receiver of claim 8, wherein the means for filtering said first and second sequences further comprises means for compensating for differences in effective group delay of said first and second sequences to produce third and fourth sequences.

13. The receiver of claim 8, wherein said plurality of weights are determined using the formula:

$$w_{k,n} = \frac{1}{filtd_{k,n} \cdot \left(1 + \left(\frac{filtd_{k,n}}{filtv_{k,n} - filtd_{k,n}}\right)^4\right)}$$

wherein $w_{k,n}$ represents said plurality of weights, $filtv_{k,n}$ represents said third sequence, and $filtd_{k,n}$ represents said fourth sequence.

14. The receiver of claim 13, wherein if $filtv_{k,n} > 1.5 \bullet filtd_{k,n}$, then $w_{k,n}$ is set to zero.

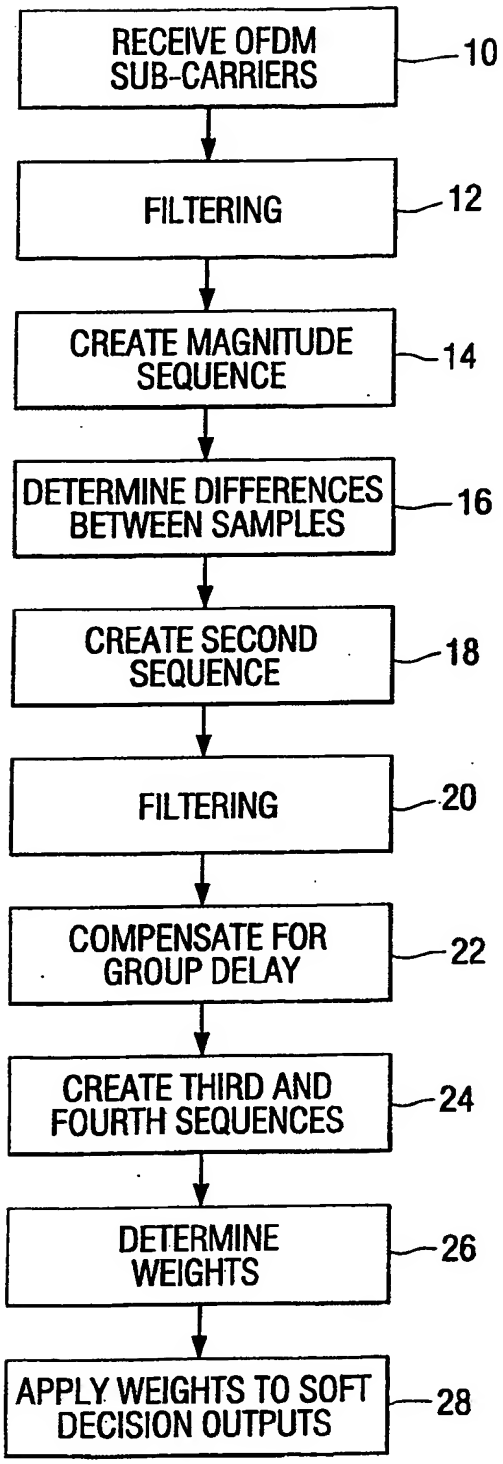
RECEIVE OFDM SUB-CARRIERS
10
FILTERING
12
CREATE MAGNITUDE SEQUENCE
14
DETERMINE DIFFERENCES BETWEEN SAMPLES
16
CREATE SECOND SEQUENCE
18
FILTERING
20
COMPENSATE FOR GROUP DELAY
22
CREATE THIRD AND FOURTH SEQUENCES
24
DETERMINE WEIGHTS
26
APPLY WEIGHTS TO SOFT DECISION OUTPUTS
28

FIG. 1

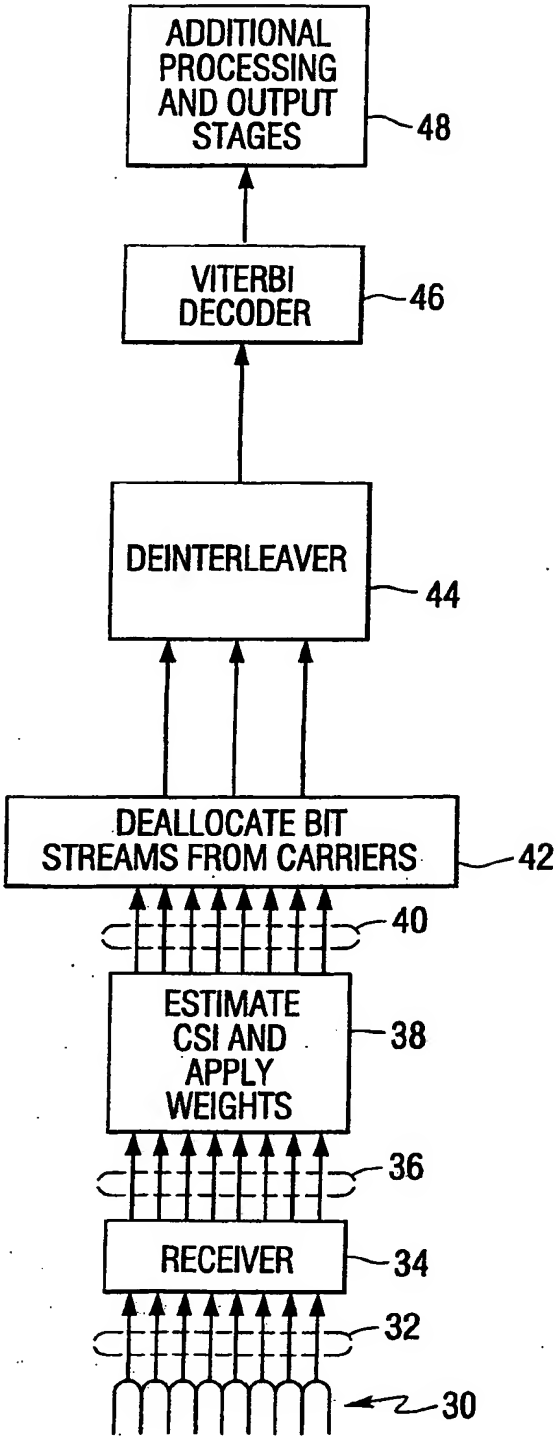
ADDITIONAL PROCESSING AND OUTPUT STAGES
48
VITERBI DECODER
46
DEINTERLEAVER
44
DEALLOCATE BIT STREAMS FROM CARRIERS
42
40
ESTIMATE CSI AND APPLY WEIGHTS
38
36
RECEIVER
34
32
30

FIG. 2

INTERNATIONAL SEARCH REPORT

Inte. onal Application No
PCT/US 99/20275

A. CLASSIFICATION OF SUBJECT MATTER
IPC 7 H04L27/26 H04L25/03 H04H1/00

According to International Patent Classification (IPC) or to both national classification and IPC

B. FIELDS SEARCHED

Minimum documentation searched (classification system followed by classification symbols)
IPC 7 H04L H04H

Documentation searched other than minimum documentation to the extent that such documents are included in the fields searched

Electronic data base consulted during the international search (name of data base and, where practical, search terms used)

C. DOCUMENTS CONSIDERED TO BE RELEVANT

Category °	Citation of document, with indication, where appropriate, of the relevant passages	Relevant to claim No.
X	BRIAN KROEGER ET AL: "Robust modem and coding techniques for FM hybrid IBOC DAB" IEEE TRANSACTIONS ON BROADCASTING, IEEE INC. NEW YORK, US. vol. 43, no. 4, December 1997 (1997-12), page 412-420 XP002104794 ISSN: 0018-9316 abstract paragraph '00II! figure 2 -----	1-14

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☐ Patent family members are listed in annex.

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Date of the actual completion of the international search	Date of mailing of the international search report
19 January 2000	26/01/2000

Name and mailing address of the ISA	Authorized officer
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